US005656908A

United States Patent [19]

Rehm

[11] **Patent Number: 5,656,908**

[45] **Date of Patent: Aug. 12, 1997**

[54] **NEGATIVE COEFFICIENT FILTER FOR USE WITH AC MOTORS**

[75] Inventor: **Thomas J. Rehm**, Mequon, Wis.

[73] Assignee: **Allen-Bradley Company, Inc.**, Milwaukee, Wis.

[21] Appl. No.: **453,297**

[22] Filed: **May 30, 1995**

[51] **Int. Cl.**[6] **G05B 1/02**

[52] **U.S. Cl.** **318/608**; 318/615; 318/632; 388/813

[58] **Field of Search** 318/609, 610, 318/561, 650, 602, 603, 632, 615, 601, 600, 608; 323/282, 283; 388/805, 813, 812

[56] **References Cited**

U.S. PATENT DOCUMENTS

5,493,192	2/1996	Nihei et al.	318/610
5,495,158	2/1996	Schmidt et al.	318/561
5,514,947	5/1996	Berg	323/282
5,545,957	8/1996	Kubo et al.	318/650

Primary Examiner—John W. Cabeca

Attorney, Agent, or Firm—Michael A. Jaskolski; John M. Miller; John J. Horn

[57] **ABSTRACT**

A method and apparatus to be used with the motor controller for reducing phase shift between the feedback signal and the correction signal. Wherein the feedback signal is compared by the controller to a reference signal to provide the correction signal used to alter motor control signals. The method includes changing a correction signal according to a transfer function to produce a filter signal wherein the transfer function includes two variables: k_n and ω_n, ω_n always being greater than zero, and k_n being negative when the feedback signal is phased shifted into a medium frequency range. The method and apparatus produce filtered signals that are phase shifted and amounts determined by the k_n and ω_n variables. Where k_n is negative, the filtered signal is phase lagged by an extreme degree.

20 Claims, 2 Drawing Sheets

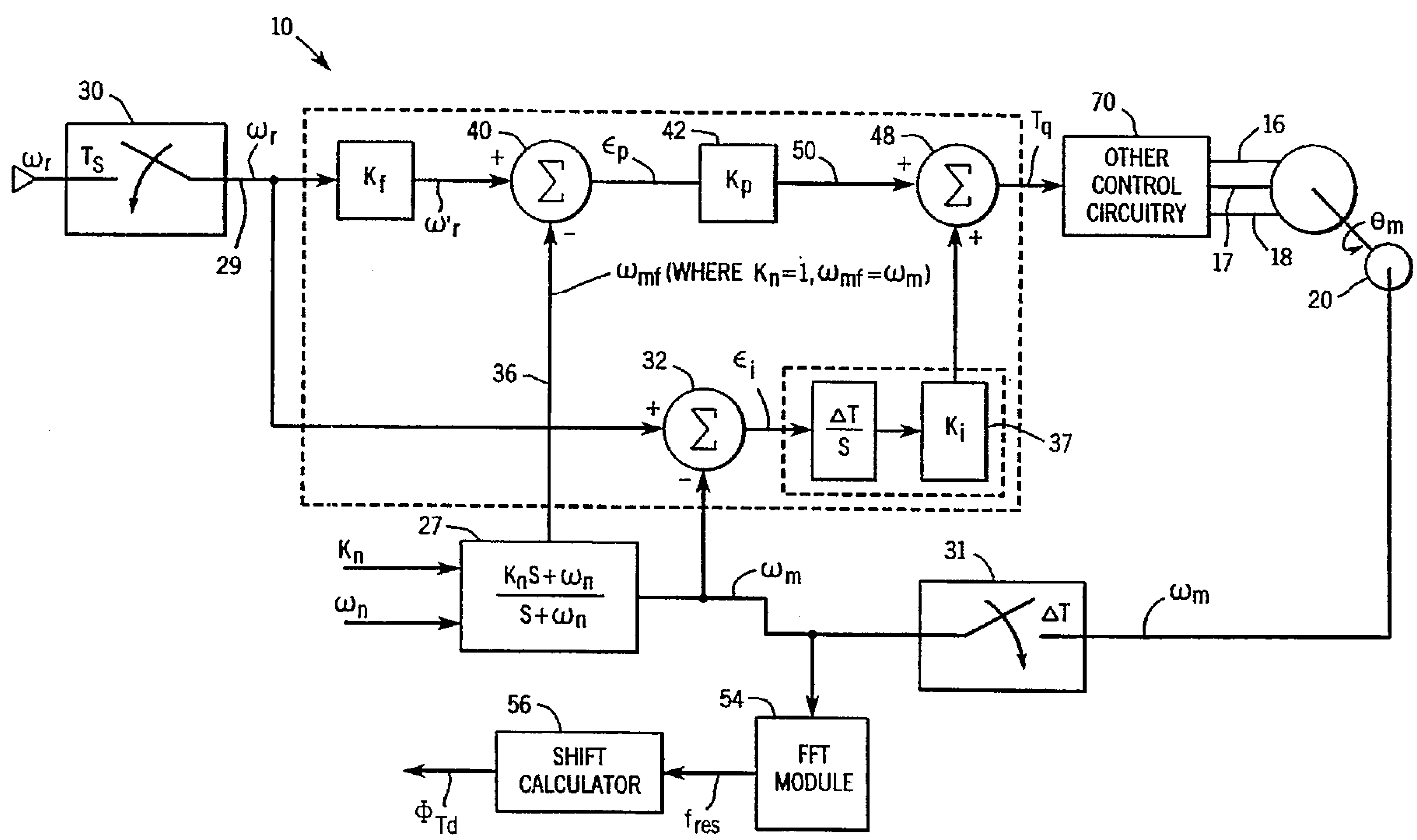


FIG. 1

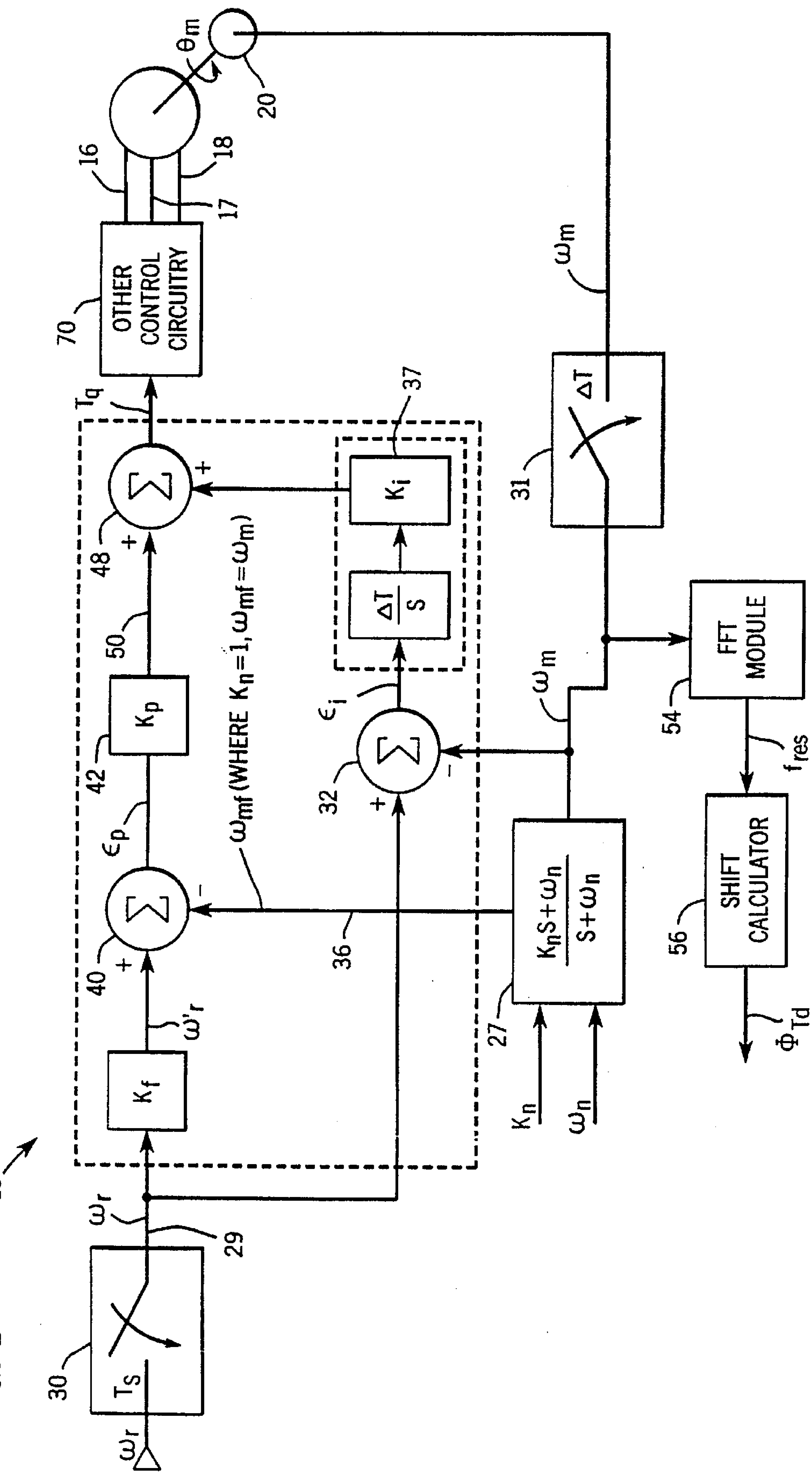

OTHER CONTROL CIRCUITRY
FFT MODULE
SHIFT CALCULATOR
$\frac{K_nS+\omega_n}{S+\omega_n}$
ω_{mf} (WHERE $K_n=1, \omega_{mf}=\omega_m$)
K_f
K_p
K_i
$\frac{\Delta T}{S}$
ΔT
T_s
ω_r
ω'_r
ω_m
θ_m
T_q
ϵ_p
ϵ_i
f_{res}
Φ_{Td}
K_n
ω_n
10
16
17
18
20
27
29
30
31
32
36
37
40
42
48
50
54
56
70

FIG. 2

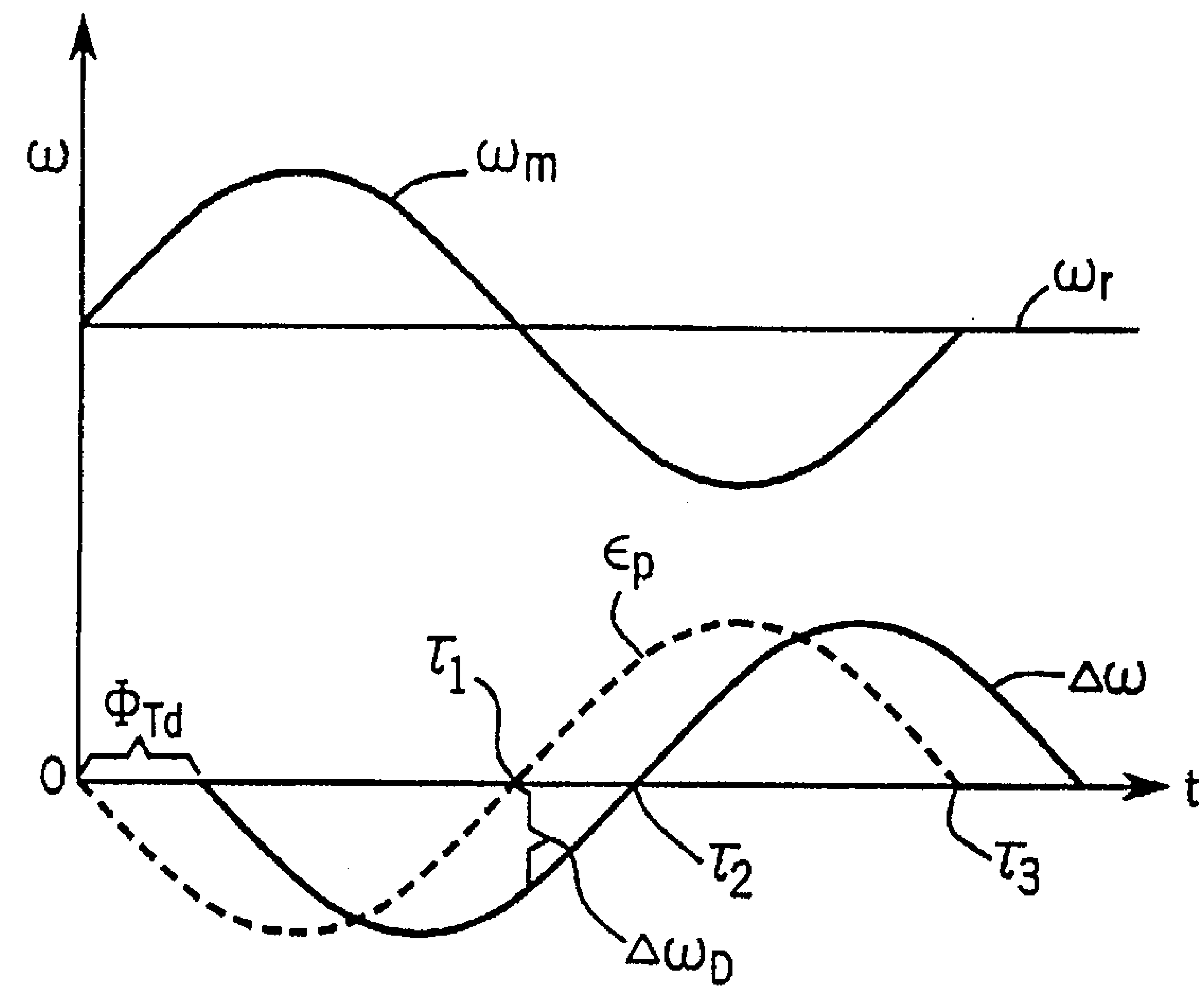


FIG. 3

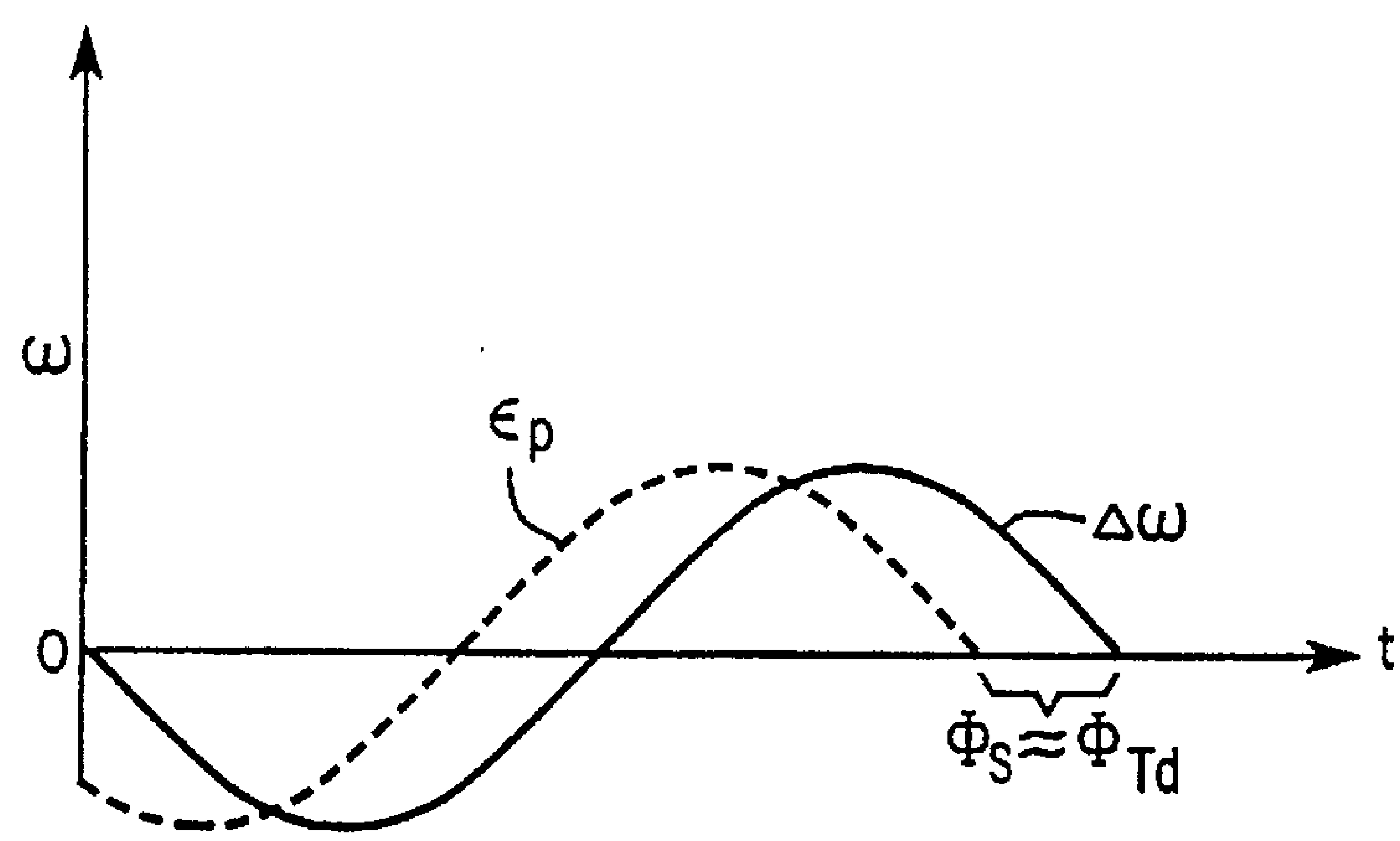
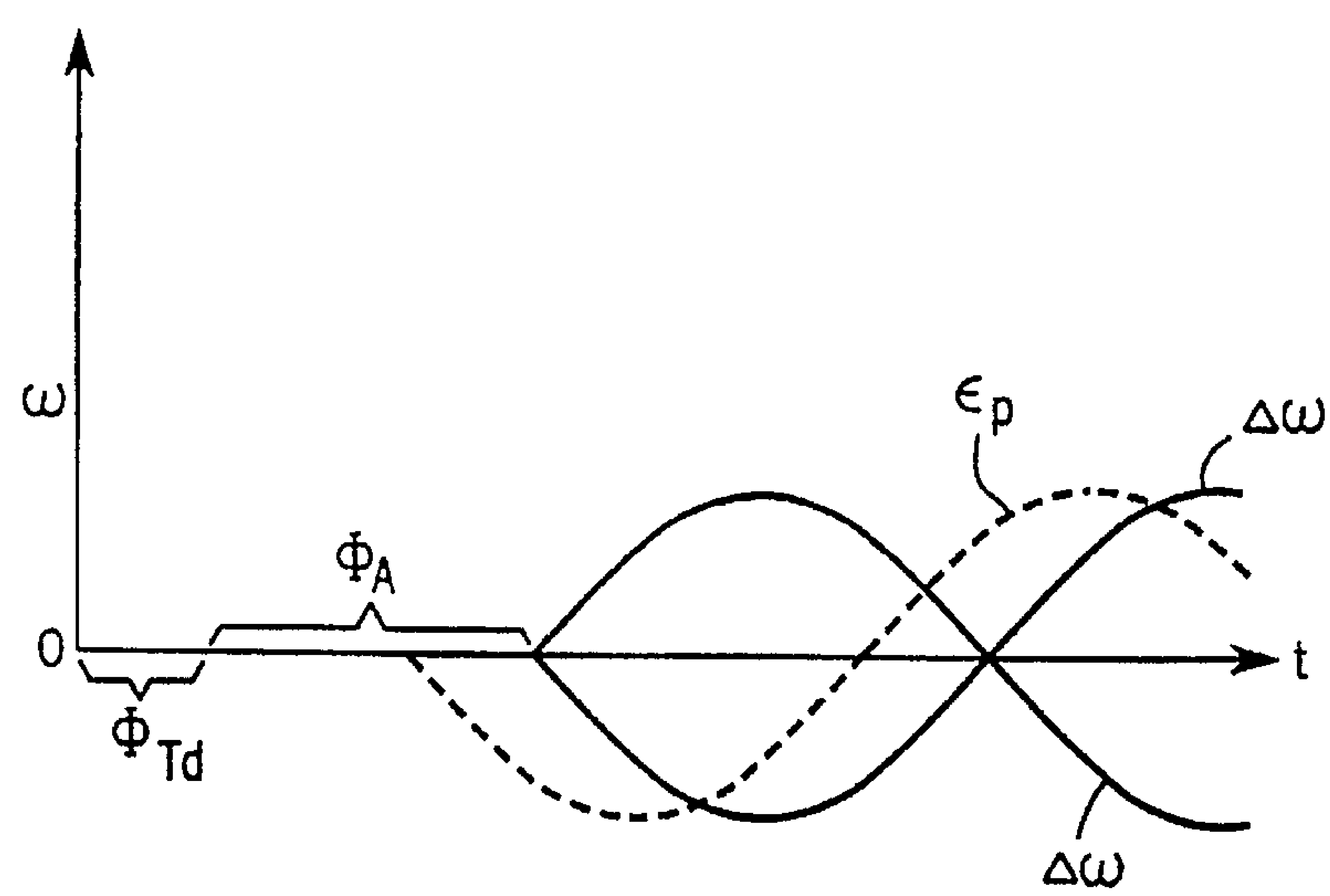


FIG. 4



NEGATIVE COEFFICIENT FILTER FOR USE WITH AC MOTORS

FIELD OF THE INVENTION

The present invention relates to systems for controlling electric motors and, more particularly, to a system for correcting for phase shift in signals propagating through a motor control system.

DESCRIPTION OF THE ART

Induction motors commonly include a rotor and a stator, the rotor positioned within a cylindrical stator frame and including a plurality of rotor windings equispaced about an external wall. An internal surface of the stator frame forms longitudinally running winding slots which receive stator windings. In this type of motor, each stator winding is connected to an AC voltage source by a separate line, the source generating an alternating current therein.

As the stator currents alternate, a wave of magnetic field flux, directed radially toward the rotor, rotates around the axis of the rotor. The relative motion between the stator flux wave and the rotor windings induces an alternating voltage and current in each of the rotor windings. The alternating rotor currents, in turn, produce a magnetic rotor flux directed outwardly toward the stator. Because of the interaction between the flux fields, the rotor encounters a force tending to rotate the rotor as the stator currents alternate. A shaft connected to the rotor rotates therewith, a distal end of the shaft being connected to a load.

Typically a motor controller is used to control motor and shaft speed. Ideally, the controller receives a reference velocity signal and produces signals which provide currents in the stator windings that impart a torque to the rotor for rotating the rotor at the reference velocity.

In reality, however, various motor parameters change during operation so that the relationship between voltages provided across stator windings and the speed at which the rotor rotates is not linear. As a result, identical voltages across the stator windings may drive the rotor at different velocities at different instances depending upon the instantaneous states of various motor parameters.

For example, as current flows through stator and rotor windings the temperature of both rotor and stator windings increases. As winding temperature changes, winding resistance changes which in turn affects the stator winding currents. This alters the stator-rotor torque relationship which finally affects shaft speed.

Because of fluctuating motor parameters, where precise shaft velocity is required, a feedback loop is often used to monitor motor velocity. The controller uses feedback information to determine how the stator winding voltages must be altered to compensate for unexpected system parameters, and then, ideally, adjusts control signals accordingly to ensure stator winding voltages that drive the motor at the desired reference velocity. Where system parameter changes occur gradually, such as stator winding resistance due to gradual temperature fluctuation, a simple feedback loop is usually sufficient. However, where system parameters change quickly at high frequencies, often a simple feedback loop can exacerbate velocity control problems as the control system cannot instantaneously alter motor velocity as a function of feedback signals. One such quickly fluctuating motor parameter is torsional shaft rotation.

Upon starting a motor, when stator and rotor fields are produced and the stator field rotates about the rotor, the rotor

experiences rotational torque. Nevertheless, the load still tends to remain stationary as the rotor begins to rotate. The shaft linking the rotor and load experiences a twisting force which effectively stresses the shaft into the direction of rotor rotation. The shaft operates as a torsional spring storing the energy of the twisting force. Once the rotor and load reach a substantially steady state speed, the energy of the twisting force is unloaded tending to sling the load around the axis of rotation to an increased velocity which is faster than the rotor velocity.

As the load is slung around, the shaft experiences a twisting force in the opposite direction, storing the energy of the force as an oppositely loaded torsional spring. This twisting force phenomenon continues to oscillate back and forth loading the shaft in one direction and then in the opposite direction. Each time the load is slung past the velocity of the rotor it imparts some velocity change to both the shaft and rotor. These velocity changes show up in the velocity feedback signal as a relatively high frequency resonating component so that the velocity feedback signal resonates about the desired reference velocity. The resonating frequency and extent of torsional shaft rotation are both functions of both motor design nuances and the load driven by the motor. Larger loads tend to have more momentum and hence spring back and forth at a slower rate (i.e., lower frequency).

Torsional shaft rotation is problematic for a number of reasons. First, this rotation results in noisy motor operation as fluctuating velocity vibrates both the load and the rotor. Second, and perhaps most importantly, as the load and shaft vibrate, the shaft often heats up to the point where the shaft and adjacent motor components may be either damaged or destroyed. Because the extend of torsional shaft rotation is a function of motor design nuances and load, it is extremely difficult to provide a solution to the problem that will work in all instances.

With feedback loops, the idea is to determine the error between desired and actual velocities and alter torque delivered to the motor in order to conform the actual velocity to the desired velocity. Hence, where the torsional force is instantaneously driving the shaft past the desired velocity in the clockwise direction, a torque in the counterclockwise direction of a magnitude equal to that of the torsional force can be provided. Unfortunately, processing delays inherent in all electronic feedback loops render simple feedback loops unable to compensate for velocity errors where high frequency resonance exists.

Most controllers include digital and/or analog control circuitry. Each controller calculation requires a finite amount of time to complete. The plurality of calculations required to generate control signals can often take a fraction of a second. This means that by the time a feedback signal is compared to a reference signal and a corrected torque signal propagates through the controller to adjust motor torque to compensate for a velocity error, the corrected motor torque is out of phase with the original velocity error (i.e., is later in time or "lags" the velocity error). Any correction based on a lagging motor torque can alter the instantaneous rotor velocity in an unintended manner, thus exacerbating the velocity error problem. While minimal lag is tolerable as the feedback loop can compensate for the majority of velocity error, excessive lag cannot be tolerated.

One way to correct for the phase shift in the correction signal is to provide delay means in the feedback loop to phase shift the corrected torque signal further. Because the resonant velocity error due to torsional shaft rotation is periodic, if the corrected motor torque is phase shifted sufficiently, the corrected motor torque corresponding to a velocity at a first instance can be used to correct a velocity error at a second instance approximately one-half period after the first instance. After additional phase shifting, if the feedback and correction signals are approximately 180° out of phase, the velocity error can be eliminated.

Delays have been provided by using analog capacitor-resistor networks in the feedback loop as known in the art. By providing adjustable value resistors, variable delay can be introduced into a circuit. While effective, the analog solution is relatively clumsy as it can require many adjustments to shift the corrected signal appropriately and such adjustments must be manually made by tweaking appropriate hardware.

A relatively effective way to correct for high frequency parameter changes such as torsional shaft rotation, is to employ an adjustable digital filter feedback loop that phase shifts the corrected motor torque in either the lagging or leading directions. By allowing filter parameters to be altered, the degree of phase shift can be changed somewhat to provide better damping characteristics. This type of system is advantageous over the analog solution because it is more versatile and parameters can be altered by way of software.

Unfortunately, adjustable digital filters have proven effective for only some degrees of phase shift while other degrees cannot be achieved without driving the controller into an unstable state. For example, most first order filters can only phase shift a correction signal or torque a maximum of approximately 40°, and that maximum can only be achieved under specific and ideal circumstances, the typical maximum being much less than 40°. Where controller phase shift is approximately 90°, even the maximum additional shift (i.e., 40°) results in a total shift of 130°, far less than the desired approximately 180°.

Where an adjustable filter cannot affect appropriate damping, analog damping systems must be specially designed to accommodate specific loads. This is extremely costly and time consuming.

Thus, it would be advantageous to have a filter that could compensate for phase lag in a feedback signal to account for phase shift due to controller signal propagation delay regardless of the degree of propagation delay.

SUMMARY OF THE INVENTION

The present invention includes a method to be used with a motor controller for reducing phase shift between a feedback signal and a correction signal. The feedback signal is compared by the controller to a reference signal to provide the correction signal used to alter motor control signals. The method comprises the steps of deriving a feedback signal, determining the degree of phase shift between the feedback signal and the correction signal, changing the correction signal according to a transfer function to produce a filtered signal, wherein the transfer function is the following function:

$$\frac{k_n s + \omega_n}{s + \omega_n} \qquad (1)$$

where k_n and ω_n are variables, ω_n is always greater than zero, and k_n is negative when the correction signal was phase shifted into a medium frequency range between a small angle that is greater than 75° and a large angle that is less than 105°.

Hence, one object of the present invention is to provide a filter compensator which can phase lag a signal by a greater degree than other conventional filter mechanisms. Whereas other filter mechanisms typically only phase lag correction signals by a maximum of X degrees under ideal circumstances prior to driving a controller into an unstable state, the filter of the present invention can, under certain circumstances, phase lag a signal by at least 2×, twice as much, without causing an unstable condition. Hence, even where controller phase lag is approximately 90°, the present filter can promote additional lag so that the resulting cumulative lag approaches 180° and the time delay and resulting velocity error can be compensated for.

Preferably, the small angle is 85° and the large angle is 95°. Most preferably, the small angle is 88° and the large angle is 92°. In one embodiment, the step of changing the correction signal includes a step of multiplying the correction signal by the transfer function. In another embodiment, the step of changing the correction signal includes a step of multiplying the feedback signal by the transfer function. Thus, the present inventor has realized that the inventive filter can be used at any point in the feedback loop, including the feed forward portion, to compensate for controller time delay.

In a preferred embodiment, if the feedback signal is phase shifted less than the small angle, the correction signal is changed according to the transfer function to produce a filtered signal wherein k_n is greater than 1. If the feedback signal is phase shifted past the large angle, the correction signal is changed to produce a filtered signal according to the transfer function wherein k_n is greater than or equal to 0 and less than or equal to 1.

Thus, another object of the present invention is to provide a filter mechanism which can add phase lag or phase lead to a correction signal to compensate for any degree of phase lag between the correction signal and the feedback signal. By equating k_n with a value greater than 1, phase lead can be provided. By equating k_n with a value greater than or equal to 0 and less than or equal to 1, a lesser degree of phase lag can be provided for instances where controller phase lag is much greater than 90°.

Also, in a preferred embodiment, where the phase lag due to controller calculation time is within the medium range, the k_n value is set equal to −1. By setting the k_n value equal to −1 and appropriately adjusting the ω_n value, the filter of the present invention operates as a transport function which phase lags the correction signal appropriately without affecting the magnitude of that correction signal.

In addition to the method described above, the present invention also contemplates an apparatus which can be used with the method for motor control.

The foregoing and other objects and advantages of the present invention will appear from the following description. In the description, references made to the accompanying drawings which form a part hereof, and in which there is shown by way of illustration and preferred embodiment of the invention. Such embodiment does not necessarily represent the full scope of the invention, however, a reference is made therefore to the claims herein for interpreting the scope of the invention.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a schematic showing a motor controller including a feedback loop according to the present invention;

FIG. 2 is a graph illustrating signals at various points in the controller of FIG. 1;

FIG. 3 is a graph similar to the graph of FIG. 2 illustrating phase lead compensation; and

FIG. 4 is a graph similar to the graph of FIG. 2 illustrating phase lag compensation.

DETAIL DESCRIPTION OF THE EMBODIMENT

Referring to FIG. 1, the present invention will be described in the context of an exemplary control controller **10** for controlling an AC induction motor **12**. The controller **10** receives a reference velocity signal ω_r and generates voltages on lines **16**, **17**, and **18** for exciting stator windings of the AC motor **12**. The voltages in the stator windings produce currents which induce magnetic fields in the stator. Each of the three voltages is sinusoidal and the voltages are out of phase by 120° so that the currents alternate and the stator magnetic field rotates about the stator. When the stator magnetic fields rotate they induce current in the rotor windings and a rotor magnetic field that interacts with the rotating stator field. The interaction between fields provides a rotating torque to the motor which tends to rotate the rotor about a rotation axis at a motor velocity approximately equal to the reference velocity ω_r. In order to ensure that the motor velocity ω_m is nearly identical to the reference velocity ω_r, a feedback loop is provided.

The feedback loop includes an encoder **20** that detects the angular displacement θ_m of the rotor and produces a motor velocity signal ω_m. The motor velocity signal ω_m passes through a digital sampler **31** and the filter **27** of the present invention, which is described in more detail below, to produce a filtered motor velocity signal ω_{mf} and digital motor velocity signal ω_m which are both provided to the controller **10**. As well known in the art, the controller **10** is a proportional-integral (PI) controller. The controller **10** uses the digital sampler and filter and velocity signals ω_m and ω_{mf} to correct for differences between the actual motor velocity and the reference velocity ω_r. At steady state, the reference velocity ω_r remains constant whereas the actual digital and filtered signals ω_m, ω_{mf} may fluctuate about the reference velocity ω_r due to changing motor parameters.

For the purpose of explaining the problem solved by the invention, it will be assumed that the value of k_n in the inventive filter **27** is 1 so that the filter transfer function reduces to 1 and the feedback loop is effectively filter free. Where this is the case, $\omega_{mf}=\omega_m$. Referring also to FIG. **2**, a typical reference velocity signal ω_r and associated motor velocity signal ω_m can be observed. In FIG. 1, a digital sampler **30** receives the reference velocity signal ω_r and produces a digital reference signal which is provided to the PI controller **10** on line **29**. When the controller **10** receives the reference and motor velocity signals ω_r, ω_m, each signal ω_r, ω_m is supplied to both legs of the PI controller **10**. In the proportional leg, the reference signal ω_r is stepped up by a constant k_f producing a stepped up reference signal ω'_r that has a value suitable for calculations (i.e., suitable for comparison to the motor velocity signal ω_m). Within the proportional leg, a summer **40** is used to subtract the motor velocity signal ω_m from the stepped up reference velocity signal ω_r to produce an error signal ϵ_p. Similarly, in the integral leg, a second summer **32** is used to subtract the digital motor velocity signal ω_m from the digital reference signal ω_r to produce an error signal ϵ_i.

The error signals ϵ_p and ϵ_i are provided to proportional and integral blocks **42**, **37** respectively, which cooperate to increase the speed at which the controller **10** can alter motor velocity to eliminate velocity errors. Operation of the proportional and integral legs of the PI controller is well known in the art and hence will not be explained here in detail. The outputs from the proportional and integral blocks **42**, **37** are added at a third summer **48** to produce a torque signal T_q which is provided to other control circuitry **70** which in turn generates voltages on lines **16**, **17**, and **18**. The voltages on lines **16**, **17**, and **18** alter torque delivered to the motor **12** and hence alter motor speed as a function of the torque signal T_q.

For the purposes of simplifying the description of this invention, a motor torque T_m will be identified as the torque actually delivered to the motor **12** as a result of the torque signal T_q. The motor torque T_m is not a directly generated value, but is rather the result of altered stator winding voltages as a function of the torque signal T_q provided by the PI controller. The motor torque T_m has a motor velocity altering effect which will be referred to herein as velocity correction $\Delta\omega$.

Upon starting a motor, a reference velocity signal ω_r is provided to the controller **10**. Because the motor is initially at rest, the motor velocity signal ω_m is initially zero and both error signals ϵ_p and ϵ_i equal the reference velocity ω_r. At start-up, the control circuitry **70** produces voltages and begins to accelerate the rotor. Over the course of a start-up period, the motor velocity ω_m increases until the motor velocity ω_m is equal to the reference velocity ω_r, the error signals ϵ_p and ϵ_i are reduced to zero, and the voltages on lines **16**, **17**, and **18** reach a steady AC state.

During operation, however, as motor parameters change due to torsional shaft rotation or the like, the motor velocity ω_m fluctuates about the reference velocity ω_r so that the error signals ϵ_p and ϵ_i continuously hover around a zero value, being periodically positive and negative. These fluctuations result from torsional energy stored in the shaft that produces a torsional force that oscillates between the clockwise and counterclockwise directions about the shaft axis.

In correcting for these motor velocity deviations, the object is to provide torque in a direction which opposes and which compensates for the torsional force. Hence, when the torsional force is rotating the shaft in a clock-wise direction, an equal and opposite force in the counter clock-wise direction can be provided. Similarly, when the torsional force is acting in the counter clock-wise direction, a compensating force is provided in the clock-wise direction.

Ideally the control system would be able to instantaneously determine motor velocity deviations, alter the reference velocity signal ω_r to compensate for the velocity deviations, and produce correcting motor torque that would counter the torsional rotation forces resulting in a motor velocity ω_m equal to the reference velocity ω_r.

In reality, however, after the motor velocity ω_m is sampled, the motor velocity signal ω_m is used in a plurality of complex calculations to generate a desired correcting torque signal T_q and then to produce line voltages that will generate a compensating and correcting motor torque T_m which results in the appropriate velocity correction $\Delta\omega$. While each calculation requires only a short period to complete, the cumulative time required to complete necessarily sequential calculations adds up to produce a delay T_d between the time at which the motor velocity is sensed and the time at which the motor torque T_m can be provided to correct for velocity deviations. The result is that compensating motor torque T_m is often out of phase with the velocity errors that give rise to the compensating motor torque T_m and hence the velocity correction $\Delta\omega$ is also out of phase. Where this is the case, compensating motor torque T_m can exacerbate motor control problems.

This phase shifting only minimally effects the integral leg of the PI controller. Because the integral leg integrates deviations over a relatively long period, the effects of the periodic deviations integrated generally cancel each other. Hence, even though the motor torque ω_M and resulting correction due to the integral leg are out of phase, this correction does not appreciably affect high frequency velocity errors.

However, the phase shift between motor velocity measurement and the velocity correction $\Delta\omega$ greatly affects output of the proportional leg which reacts to instantaneous high frequency velocity changes and hence greatly affects motor operation.

Referring still to FIG. 2, because of processing time required to produce the error signal ϵ_p, the error signal ϵ_p itself would probably be out of phase slightly. Nevertheless, for the purpose of this description, it will be assumed that the error signal ϵ_p is in phase with the motor velocity. However, because of controller time delay T_d, the velocity correction $\Delta\omega$ lags the motor velocity signal ω_m by a controller phase angle Φ_{Td}. Where this is the case, there will be times when the motor torque T_m and resulting velocity correction $\Delta\omega$ tend to increase motor velocity where it is desired to decrease the velocity and vice versa.

For example, at τ_1, where the error signal ϵ_p is equal to zero, because of the phase lag Φ_{Td}, instead of generating zero motor torque T_m and zero velocity correction $\Delta\omega$ as desired, the velocity correction $\Delta\omega$ will be $\Delta\omega_D$ which tends to slow the motor. Between times τ_1 and τ_2, the motor torque T_m will continue to slow the motor even though, referring to the motor velocity signal ω_m, clearly a motor torque T_m and velocity correction $\Delta\omega$ tending to increase motor velocity ω_m is desired. Between times τ_2 and τ_3, the velocity correction $\Delta\omega$ is in the correct direction, tending to speed up the motor, however, it is of an imperfect amplitude because it is not symmetrical with the second portion of the motor velocity signal ω_m.

When the motor velocity ω_m and velocity correction $\Delta\omega$ are out of phase, the velocity correction $\Delta\omega$ will partially correct for velocity errors during some portion of each one-half cycle of the motor velocity ω_m (i.e. between τ_2 and τ_3) and will exacerbate errors during the other portion (i.e. between τ_1 and τ_2). The degree to which the velocity correction $\Delta\omega$ corrects is a function of the extent of phase lag between the motor velocity signal ω_m and the velocity correction $\Delta\omega$, and hence is a function of the controller time delay T_d.

Where the motor velocity signal ω_m is characterized by a relatively long period, a finite controller delay period T_d may be relatively minimal. Where this is the case it may be desirable to use an unaltered motor torque T_m to change motor velocity as the adverse effects of the delay will be relatively minimal. However, where the motor velocity signal ω_m is characterized by a short period, a finite controller delay period T_d will be relatively large resulting in a large phase shift Φ_{Td} so that a velocity correction $\Delta\omega$ adversely effects motor operation.

Because of the sequential nature of calculations required to generate the motor torque T_m and produce a velocity correction $\Delta\omega$, the time delay T_d for a specific controller cannot be appreciably altered. However, where the time delay T_d for a specific controller is known and the frequency of the resonating motor velocity signal can be determined, the phase shift Φ_{Td} can be determined and additional phase shift, either leading or lagging, can be provided. In this manner, the motor torque T_m and hence velocity correction $\Delta\omega$ can be placed more in phase with the motor velocity signal ω_m to achieve better motor control.

Referring still to FIG. 2, the controller phase lag is Φ_{Td}. Referring also to FIG. 3, by adding phase lead Φ_s equal to the phase lag Φ_{Td} to the motor torque T_m, the resulting velocity correction $\Delta\omega$ can be made more in phase with the motor velocity signal ω_m and hence can correct for velocity deviations.

Another solution to the phase lag problem is to phase lag the velocity correction $\Delta\omega$ so that when the additional phase lag Φ_A is added to the controller phase lag Φ_{Td}, the total lag approaches 180°. In FIG. 4, the velocity correction $\Delta\omega$ is phase lagged by Φ_A where Φ_A is approximately $180°-\Phi_{Td}$. If the resulting velocity correction $\Delta\omega$ (which is approximately 180° out of phase with the motor velocity signal ω_m) is inverted to produce inverted velocity correction $\Delta\omega'$, the velocity correction $\Delta\omega'$ is in phase with the motor velocity signal ω_m.

Unfortunately, often when a signal is phase lagged using a digital filter, the resulting signal amplitude is also affected. In other words, digital filters generally do not operate as pure transport mechanisms in either the phase lag or lead directions. The degree of amplitude alteration is typically a function of the extent of phase lag or lead produced by the filter. Hence, under certain circumstances, for example, where controller produced lag is minimal (i.e.<20%), the velocity correction $\Delta\omega$ is sufficiently in phase with the motor velocity signal that it can correct for velocity fluctuations and therefore no filtering is required.

Where phase lag due to a time delay T_d is substantial yet relatively minimal (i.e. between 20° and 85°), phase lead techniques are preferable to phase lag because the signals need only be shifted through a relatively small angle in order to achieve acceptable control. For example, where controller phase lag is approximately 60 degrees, by adding phase lead of 30 degrees, the velocity correction $\Delta\omega$ can be brought into an acceptable phase relationship with the motor velocity ω_m without substantially altering the amplitude of the velocity correction $\Delta\omega$. Similarly, where phase lag is relatively large (i.e., between 95° and 160°), phase lag techniques are preferable.

Referring again to FIGS. 1 and 2, in order to determine the degree of phase lag Φ_{Td}, and hence how a velocity correction $\Delta\omega$ must be shifted to correct for velocity errors, a resonating frequency f_{res} of the motor velocity signal ω_m must be determined. To this end, the digital motor velocity signal ω_m can be provided to a fast fourier transform (FFT) module **54** which performs an FFT on the digital signal ω_m to produce the resonating frequency signal f_{res}. The phase lag can be determined from the resonating frequency f_{res} and the time delay T_d according to the following equation:

$$\Phi_{Td}=2\pi f_{res}\cdot T_d=\omega_{res}\cdot T_d \tag{2}$$

where Φ_{Td} is the phase shift due to the controller time delay T_d and ω_{res} is the resonating velocity in radians. From equation 2 it can be observed that the phase shift Φ_{Td} increases where either the time delay T_d or the resonating frequency f_{res} increases. The frequency signal f_{res} is supplied to a shift calculator **56** which calculates the phase shift Φ_{Td} between the motor velocity signal ω_m and the velocity correction $\Delta\omega$ according to Equation 2. The phase shift Φ_{Td} is output to an operator by a digital readout or the like (not shown). Once controller phase lag Φ_{Td} is determined, an operator can determine how the motor torque T_m and velocity correction should be shifted according to the principles identified above.

Referring again to FIG. 1, the filter according to the present invention implements the transfer function of Equation 1 where k_n and ω_n are constants, ω_n is always positive, and k_n can be positive, zero, or negative, depending upon the required lag or lead compensation. Both k_n and ω_n are input by an operator using a keypad or the like (not shown) as a function of both phase shift Φ_{Td} and the additional phase shift Φ_A required to shift the velocity correction $\Delta\omega$ into acceptable orientation with respect to the motor velocity ω_m.

Given the filter transfer function as expressed in Equation 2, it has been determined that there are four different categories of values that can be used for k_n, each of which affects phase shift in a different manner.

The effect of k_n

Referring to again Equation 1, by dividing the numerator by the denominator to provide the first three terms in a series, equation 1 can be represented as:

$$1+\frac{(k_n-1)}{\omega_n}s-\frac{(k_n-1)}{\omega_n^2}s^2+\ldots \tag{3}$$

If a time delay is small, it is possible to approximate the time delay transfer function by the truncated power series:

$$e^{-sT_d}=1-sT_{dA}+\frac{S^2T_{dA}^2}{2}-\ldots \tag{4}$$

Equations 3 and 4 are of similar forms, but the third terms are different and hence the two equations are not completely comparable. However, because the first two terms in each equation dominate the solution of each equation, what is known about one equation can be used to understand some general characteristics of the other equation. For example, by setting the time delay T_{dA} from Equation 4 equal to the similar term in Equation 3, Equation 5 can be derived:

$$T_{dA}=\frac{1-k_n}{\omega_n} \tag{5}$$

Equation 5 can be used to establish how different k_n values generally affect phase shift and time delay.

For example, where k_n is greater than 1, the added time delay T_{dA} will be negative, despite the magnitude of ω_n. An negative time delay T_{dA} indicates phase lead added to a system. Similarly, where k_n is greater than or equal to zero and less than 1, the time delay T_{dA} will be positive indicating that a phase lag is being added to the system. Where k_n is 1, the time delay T_{dA} is zero and hence, at this limiting condition, no phase shift is introduced into the system.

Importantly, for the purposes of the present invention, it has been determined that extreme phase lag, greater than that which can be realized by using other filter systems, can be generated by setting k_n equal to a negative value. For example, the greatest amount of phase lag that can be generated using any non-negative value of k_n is $1/\omega_n$ where k_n is equal to zero (this is a limiting case at which ω_n is defined to be a low pass filter bandwidth). By setting k_n equal to negative 2, the time delay lag T_{dA} can potentially be increased three fold to $3/\omega_n$. By setting k_n equal to negative 4, the time delay lag T_{dA} can potentially be increased five fold to $5/\omega_n$.

In addition, while negative k_n values result in large phase lags, phase lags produced with negative k_n values alter velocity correction $\Delta\omega$ amplitudes relatively less than lags provided with positive k_n values. Hence, more lag can be introduced with less amplitude variation.

It should be noted that these figures correspond to Equation 5 which, as already indicated, is not entirely accurate but is only an approximation of the time delay T_{dA}. Nevertheless, these figures give some indication of what can be expected by setting k_n equal to certain values. In addition, these forms generally conform to experiments wherein different values of k_n have been used to alter phase shift.

Referring again to Equation 3, in the special circumstance where k_n is equal to negative 1, at least the first three terms of Equations 3 and 4 assume identical form. In this case, setting k_n equal to negative 1 in Equation 5, an extremely precise value of time delay T_{dA} can be expressed as:

$$T_{dA}=2/\omega_n \tag{6}$$

Where a large phase lag must be added to a system k_n can be set equal to negative 1 and ω_n can be adjusted accordingly to provide desired phase lag. Where kn=−1, the transfer function of Equation 1 represents a pure time transport and very little amplitude variation takes place as a result of the lag.

In situations calling for additional phase lag, the object is to add an amount of phase lag Φ_A to the system that approximately compliments the controller produced phase lag Φ_{Td} (i.e. the phase shift Φ_{rd} from the time delay T_d and added shift Φ_A due to the added time delay T_{dA} should be approximately equal to or greater than π radians or 180 degrees). The amount of added phase lag Φ_A resulting from added time delay T_{dA} can be determined according to the following equation:

$$\Phi_A=2\pi f_{res}\cdot T_{DA}=\omega_{res}\cdot T_{dA} \tag{7}$$

Combining Equations 6 and 7 and solving for ω_n:

$$\omega_n=\frac{2\omega_{res}}{\Phi_A} \tag{8}$$

Because the phase shifts should total approximately 180 degrees, Φ_A is equal to (28.64 rad. $-\Phi_{Td}$) and therefore:

$$\omega_n=\frac{2\ \omega_{res}}{\pi\text{ rad.}-\Phi_{Td}} \tag{9}$$

Combining Equations 2 and 9 and simplifying:

$$\omega_n=\frac{2\ \omega_{res}}{(\pi\text{ rad.}-T_d\omega_{res})} \tag{10}$$

Thus, where Φ_{Td} from Equation 2 is within the problem zone between 85 and 95 degrees, k_n can be set to negative one and an approximate value of ω_n can be determined from Equation 10 as ω_{res} and the controller time delay T_d are both known. While ω_n from Equation 10 should be substantially correct, due to some inevitable amplitude variation in the velocity correction $\Delta\omega$ caused by the phase shift, the value of ω_n may have to be fine tuned to achieve the best possible balance between additional phase shift Φ_A and amplitude variation. This can easily be achieved by the controller Operator.

While the above description details various elements of an apparatus including the filter according to the present invention, it should be understood that all of the elements, including the filter, are meant to be implemented in softwares, computer programs, and represent algorithms for execution by conventional-type digital processors adapted for industrial applications such as a model 8196 microelectronic processor that is supplied by INTEL Corporation of Santa Clara, Calif.

It should be understood that the methods and apparatuses described above are only exemplary and do not limit the scope of the invention, and that various modifications could be made by those skilled in the art that would fall under the scope of the invention. To apprise the public of the scope of this invention, we make the following claims:

I claim:

1. A method to be used with a motor controller for reducing phase shift between a feedback signal and a

correction signal, the feedback signal compared by the controller to a reference signal to provide the correction signal used to alter motor control signals, the method comprising the steps of:

deriving a feedback signal;

determining the degree of phase shift between the feedback signal and the correction signal;

changing the correction signal according to a transfer function to produce a filtered signal, wherein the transfer function is the following function:

$$\frac{k_n s + \omega_n}{s + \omega_n}$$

where k_n is a filter gain coefficient and ω_n is a filter cutoff frequency, ω_n is always greater than zero, and k_n is negative when the feedback signal is phase shifted into a medium frequency range between a small angle that is greater than 75 degrees and a large angle that is less than 105 degrees.

2. The method as recited in claim **1** wherein the small angle is 85 degrees and the large angle is 95 degrees.

3. The method as recited in claim **1** wherein the small angle is 88 degrees and the large angle is 92 degrees.

4. The method as recited in claim **1** wherein the step of changing the correction signal includes the step of multiplying the correction signal by the transfer function.

5. The method as recited in claim **1** wherein the step of changing the correction signal includes the step of multiplying the feedback signal by the transfer function.

6. The method as recited in claim **1** further including the steps of:

if the feedback signal is phase shifted less than the small angle, changing the correction signal according to the transfer function to produce a filtered signal wherein k_n is greater than one; and

if the feedback signal is phase shifted past the large angle, changing the correction signal to produce a filtered signal according to the transfer function wherein k_n greater than or equal to zero and less than or equal to one.

7. The method as recited in claim **1** wherein the controller introduces a known time delay into the correction signal and a periodic resonating signal is included in the feedback signal and the step of determining a phase shift includes the steps of:

(a) determining the frequency of the periodic resonating signal; and

(b) mathematically combining the periodic resonating signal and the time delay signal to determine the phase shift.

8. The method as recited in claim **7** wherein the step of mathematically combining includes the step of multiplying the periodic resonating signal by the time delay signal to determine the phase shift in radians.

9. The method as recited in claim **1** wherein k_n is negative 1.

10. The method as recited in claim **7** wherein k_n is negative and:

$$\omega_n \leqq \frac{2\,\omega_{res}}{\pi\ \text{rad.} - T_d\omega_{res}}$$

where ω_{res} is the resonating frequency in radians and T_d is a time delay introduced into the correction signal by the controller.

11. An apparatus to be used with a motor controller for reducing phase shift between a feedback signal and a correction signal, the feedback signal compared by the controller to a reference signal to provide the correction signal used to adjust motor control signals, the apparatus comprising:

a sensor for deriving a feedback signal;

a calculator for determining the degree of phase shift between the feedback signal and the correction signal;

a filter changing the correction signal according to a transfer function to produce a filtered signal, wherein the transfer function is the following function:

$$\frac{k_n s + \omega_n}{s + \omega_n}$$

where k_n is a filter gain coefficient, and ω_n is a filter cutoff frequency provided to the filter and wherein ω_n is always greater than zero and k_n is less than zero when the feedback signal is phase shifted into a medium frequency range, the medium frequency range between a small angle that is greater than 75 degrees and a large angle that is less than 105 degrees, the filter producing a filtered signal; and

an invertor for inverting the filtered signal to produce an inverted signal.

12. The apparatus as recited in claim **11** wherein the small angle is 85 degrees and the large angle is 95 degrees.

13. The apparatus as recited in claim **11** wherein the small angle is 88 degrees and the large angle is 92 degrees.

14. The apparatus as recited in claim **11** wherein the filter includes a multiplier that changes the correction signal by multiplying the correction signal by the transfer function.

15. The apparatus as recited in claim **11** wherein the filter includes a multiplier that changes the correction signal by multiplying the filter signal by the transfer function.

16. The apparatus as recited in claim **11** wherein:

if the feedback Signal is phase shifted less than the small angle, the filter changes the correction signal according to the transfer function to produce a filtered signal wherein k_n is greater than one; and

if the feedback signal is phase shifted past the large angle, the filter changes the correction signal to produce a filtered signal according to the transfer function wherein k_n greater than or equal to zero and less than or equal to one.

17. The apparatus as recited in claim **11** wherein the controller introduces a known time delay into the correction signal and a periodic resonating signal is included in the feedback signal and the calculator includes a frequency sensor to determine the frequency of the periodic resonating signal and a combining means for mathematically combining the resonating signal frequency and the time delay signal to determine the phase shift.

18. The apparatus as recited in claim **17** wherein the combining means includes a multiplier for multiplying the resonating signal frequency by the time delay signal to determine the phase shift in radians.

19. The apparatus as recited in claim **11** wherein k_n is negative 1.

20. The apparatus as recited in claim **17** wherein:

$$\omega_n \leqq \frac{2\,\omega_{res}}{\pi\ \text{rad.} - T_d\omega_{res}}$$

where ω_{res} is the resonating frequency in radians and Td is a time delay introduced into the correction signal by the controller.

* * * * *